Friedrich Krug, DJ3RV

A Receiver for the VLF Time and Frequency Standard Transmissions from DCF 77

This receiver can be used to synchronize a phase-locked crystal oscillator for use as frequency standard for amateur applications. The requirement of an accurate time and frequency standard for the realization of modern communication technology has been mentioned several times in VHF-COMMUNICATIONS. Especially in coherent telegraphy (CCW) or similar modern communications (such as spread spectrum technology), one requires an exact frequency and time base for the time and computer-controlled digital filters and evaluation circuits. The described receiver is designed for the reception of the VLF time and frequency standard DCF 77, which transmits at a frequency of 77.5 kHz. This transmitter can be received throughout Europe.

The time information is provided by the transmitter in the form of pulse-width modulated second pulses; and the extremely constant transmit frequency is used as reference frequency to synchronize a crystal oscillator with the aid of a phase-locked loop to form a frequency standard.

The applications of such a time and frequency standard can be made very versatile by adding additional modules. The most often used application is, of course, to evaluate the time information to form a very accurate clock. However, the extremely constant frequency is more important to us radio amateurs who can use it as frequency standard.

Examples for this are some modules realized by the author to improve the frequency stability of the VHF-oscillator with DAFC described by DJ7VY in Edition 2/1981 of VHF COMMUNICATIONS, in conjunction with the receiver concept described by the author in Edition 4/1981, as well as the 96 MHz crystal oscillator for UHF/SHF local oscillators described by DK1AG also in that edition.

The author has been using a 10 MHz frequency standard, which is locked to the DCF 77 receiver, for testing, and also as a clock for frequency counters.

The actual reason for developing this receiver was to obtain a time and frequency standard for a computer-controlled RTTY transceiver, in which all frequency and time signals are coupled together so that coherent communication will be possible. This requires a circuit that is designed for high reliability.

In the last ten years, a multitude of publications and construction articles have appeared regarding receivers for the time and frequency standard transmitter DCF 77. A certain number were constructed and operated with success by the author. However, the author was forced to carry out his own development since the operation of the available receivers was not possible in the vicinity of a TV-receiver, or computer with video display. It was found that in most cases an unlock condition of the phase-locked loop in the crystal oscillator circuit, or the time control circuit was caused by

interference originating in the direct vicinity of the receiver that is mostly from one's own electrical equipment (man-made noise).

This led very quickly to a clean-up of the "electronic pollution", however, it was not possible to suppress all interference sources, which meant that a receive system was required that had a higher interference immunity.

1. RECEIVER CONCEPT

It is necessary to know the field strength to be expected and the interference field strength before conceiving the receiver. In a publication by Becker und Rohbeck (1) of the Federal Physical Laboratories (PTB) of Brunswick/ W. Germany, who generate the frequency and time signals for the DCF 77 transmitter, details are given about the usable field strength, which are based on measurements of the German PTT, as well as calculations of the propagation of the ground wave. According to this, an electrical field strength of 14.5 mV/m at a distance of 100 km, 2.3 mV/m at 400 km, and still 0.6 mV/m at 800 km is to be expected.

Of course, these values can be considerably less in the case of unfavourable locations, or where screening occurs. For instance, field strengths of between 0.6 mV/m and 8 mV/m were measured at the author's desk in a reinforced concrete building at a distance of approximately 150 km from the transmitter.

It was found that it was very much more difficult to determine the field strength of the interference. It was necessary to localize the interference sources, after which the measurement was not possible in most cases, or only with considerable expense of measuring equipment and time.

The interference sources can be split into three different groups.

A) Wideband Single-Pulse Interferences

These mainly originate from switching processes in electrical household equipment such as washing machines, electric cookers, or heaters. Most of this equipment is, unfortunately, still not switched during the zero pass of the power-line voltage using electronic components. Any radio amateur who has his antenna in the vicinity of an old lift system will know the problems encountered with this type of interference.

Thunderstorms also fall within this category of interference, as do discharge processes in non-grounded transmit antennas, and static discharge in the case of synthetic flooring and chairs. Even in the author's laboratory, a thermostat-controlled low-voltage soldering iron and the ignition processes of the fluorescent lamps caused strong spikes in the signal from the antenna.

Most of these interference sources cannot be suppressed. Any interference that is fed into the equipment by the power line, can be suppressed sufficiently well by using power-line filters in the form of lowpass filters, and by using a voltage stabilizer. Battery operation of the circuit is not possible due to the high current drain.

The interference spectrum received via the receive antenna must be suppressed as well as possible using narrow-band filters whose "ringing" is as low as possible, in order to ensure that the following circuits are not overdriven. Any residual phase jumps must be suppressed with the aid of the time constant of the phase-lock loop.

B) Low-Frequency Periodic Pulse Interferences

This is the largest category of interference sources. Frequency spectrums are generated by phase-control circuits of small electrical equipment, collector motors, vertical and horizontal deflection circuits of TV-receivers and monitors, switching power supplies, up to the data currents in computers; in the most favourable cases, this is in the form of a noise spectrum; mostly, however, they are in the form of descrete spectral lines of relatively high power density. The fifth harmonic of the line frequency of TV-receivers is extremely strong:

$$f = 5 \times 15625 \text{ Hz} = 78125 \text{ Hz},$$

which means that it is only 625 Hz above the transmit frequency of DCF 77. In the case of a portable TV-receiver owned by the author, the interference field strength of the fifth harmonic amounted to $\hat{E}$ = 300 mV/m! at a spacing of 3 m. Since the line frequency used for displaying the data of personal computers is only approximately 15625 Hz, the interference from such sources can be far nearer to the DCF-frequency.

C) Descrete Frequency Interference

This interference is usually caused in the shack itself, which means that it can very often be suppressed. Radiating shortwave feeders, and VHF-portable equipment should not be operated in the vicinity of the DCF-receive antenna.

As precautionary measure, one should provide a low-pass filter between antenna and antenna preamplifier.

Together with the knowledge of the interference effects and experience made by constructing the circuits published in (1), the concept was established for building a receive system for a simple frequency standard.

Of course, even a simple frequency standard should exhibit a good stability. For this reason, the long-term stability was determined over a period of 100 days (1), and was found to be approximately 2 x 10^{-13}, which was given by the signal of the DCF 77 transmitter.

The lower short-term stability of the DCF 77 signal caused by propagation and interference, requires a short-term stable crystal oscillator that is controlled via a PLL having a large time constant.

A frequency of f_0 = 3.1 MHz was selected as oscillator frequency. This is the 40th harmonic of the transmit frequency of DCF 77. If the crystal frequency is divided by 31, a decadic frequency of 100 kHz will result, from which all further signals are derived.

Of course, other oscillator frequencies are possible, however, require an even multiple of the DCF-frequency. This results in several useful reference frequencies, according to whether a harmonic is required that can be divided decadically, hexadecimally, or binary.

Table 1 shows the frequencies divided into primary factors, for which the PC-board layout can be used so that it is possible to also employ a crystal having another frequency.

77.5 kHz	=	$2^2 \times 5^4 \times 31$ Hz	=	f_{DCF}
3.1 MHz	=	$2^5 \times 5^5 \times 31$ Hz	=	40 x f_{DCF}
2.79 MHz	=	$2^4 \times 3^2 \times 5^4 \times 31$ Hz	=	36 x f_{DCF}
3.72 MHz	=	$2^6 \times 3 \times 5^4 \times 31$ Hz	=	48 x f_{DCF}
4.96 MHz	=	$2^8 \times 5^4 \times 31$ Hz	=	64 x f_{DCF}

Table 1: Divided multiples of the DCF-frequency

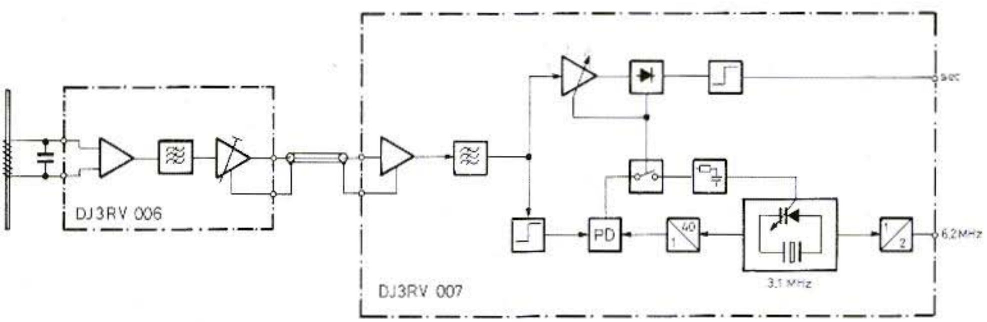

Fig. 1: Block diagram of the DCF 77 receiver

If the oscillator frequency is selected as clock for a computer – microprocessors such as the 8085 or Z 80 require clock frequencies between 2 MHz and 8 MHz – one will ensure that the signals and evaluation processes generated in the computer are directly coupled to the standard frequency.

Figure 1 shows the block diagram of the receiver concept. The signal is received using a ferrite or loop antenna. The signal is then fed to a preamplifier equipped with a narrow-band filter for suppressing the harmonics of the line frequency and is followed by an impedance converter. The antenna module DJ3RV 006 can be accommodated at a low-interference position and connected to receiver DJ3RV 007 with the aid of conventional coaxial cable of any required length (up to 100 m RG-58/U).

The receiver DJ3RV 007 possesses an impedance converter at the input which is followed by a further filter for improving the ultimate selectivity. After this, the signal is split up and fed firstly to a controlled amplifier with demodulator. This provides the time-information, and is used simultaneously for indicating a missing transmit signal. The second signal path of the 77.5 kHz signal is fed via a limiter to the phase discriminator PD. The phase comparison is made to the signal derived by dividing the oscillator frequency by 40. The phase deviation provides a DC-voltage that is passed via a low-pass filter having a large time constant, which is then used for synchronizing the oscillator frequency. If the transmitter is inoperative for any reason, it is possible for the control circuit to be disconnected.

In order to ensure a low-reactive output coupling, the oscillator signal is fed out via a frequency doubler as 6.2 MHz signal.

This concept was accommodated on two PC-boards and the author's prototype has been running for more than a year in virtually continuous operation.

2. THE ANTENNA

The first experiments with DCF-reception were made by the author approximately 10 years ago using a ferrite antenna from an old portable radio. The long-wave coil on this ferrite rod was resonated to 77.5 kHz by adding parallel capacitors, and used together with an antenna preamplifier as described by (1). This antenna had a measured bandwith of 2.6 kHz together with the relatively large parallel capacitance, which results in a Q of approximately 30. As later experiments with various antennas showed, this antenna represented a good compromise between receive energy and sensitivity to interference (purely by chance), and was only exceeded by using loop antennas.

The disadvantage of ferrite antennas is the non-linear behaviour of ferrite materials. This causes a phase modulation of the input signal in the case of strong magnetic field interferences from welding transformers, chokes in fluorescent lamps, deflection coils on TV-tubes, and the chokes in switching power supplies. The higher the Q of the antenna, the greater is the phase deviation. The interference compensates for itself statistically due to the long-term evaluation of the low-pass filter, however, it can cause a phase jump of the phase-control circuit in the case of additional interference pulses.

2.1. Calculation of the Antenna Voltage

For space reasons, only resonant loops or ferrite antennas can be used for reception of the VLF DCF 77 transmissions. The dimensions of these antennas are very small in comparison to the wavelength. Such antennas mainly utilize the magnetic field components of the signal to be received, and the directional characteristics of the antenna correspond to that of a short dipole, which is a "8" with a flat maximum and sharp minima. The output voltage of the antenna described in (3) amounts in the following, when optimumly aligned to the magnetic field component H:

$$U_{eff} = \frac{\pi \times f}{\sqrt{2}} \times A \times N \times Q \times \mu_{eff} \times \mu_0 \times H$$

and to the following, due to the combination of the magnetic field component H with the electrical field component E with the aid of the field impedance characteristic Z_0 of free space:

$$H = \frac{1}{Z_0} E = \sqrt{\frac{\varepsilon_0}{\mu_0}} \times E$$

The following is valid:

$$U_{eff} = \frac{\pi}{\sqrt{2}\,\lambda_0} \times A \times N \times Q \times \mu_{eff} \times E$$

where:

- λ_0 – Wavelength of the receive signal
- A – Loop surface, or cross section of the ferrite surface
- N – Number of turns
- Q – Operating Q of the antenna system
- μ_{eff} – Effective permeability number of the ferrite material

In the case of the loop antenna, $\mu_{eff} = 1$.

The output voltage of the antenna should be as high as possible in order to ensure good reception.

As can be seen in the equation, the voltage at the output of the antenna is directly proportional to the surface A, the number of turns N, and the Q, and to the effective permeability μ_{eff} in the case of a ferrite antenna, for any given field strength at the reception location. Unfortunately, these magnitudes cannot be increased infinitely.

The loop surface A is limited, if the antenna is to remain within handy dimensions, and is given by the cross section of the ferrite rod in the case of a ferrite antenna.

The number of turns N is limited by the stray capacity C_s of the coil. The intrinsic resonance of the coil should not be less than the receive frequency.

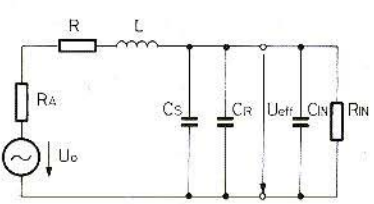

Fig. 2: Equivalent diagram of the receive antenna

The operating Q of the antenna is dependent on several magnitudes, that must be considered further in order to approximate their effects.

Figure 2 gives the equivalent diagram of the receive antenna; U_0 is the source voltage resulting from the field strength at the receive location. If the Q is not too small ($Q > 10$), the resonant frequency of the antenna is determined by the inductance L and the stray capacitance C_s of the coil, the input capacitance C_{in} of the antenna preamplifier, and the resonant circuit capacitance C_R.

$$f = \frac{1}{2\pi \sqrt{L \times (C_S + C_{in} + C_R)}}$$

The operating Q can be calculated as follows:

$$\frac{1}{Q} = \frac{2\pi\, f \times L}{R_{in}} + \frac{(R + R_A)}{2\pi\, f \times L} + \left(\frac{1}{Q_M} \times \frac{3\, l'}{4\, l}\right)$$

As can be seen in the equation, the input impedance R_{in} of the antenna amplifier connected in parallel to the resonant circuit, the loss impedance R of the wire in series with the coil, and the radiation resistance R_A determine the Q of the inductance.

In the case of ferrite antennas, one must also take the Q of the material into consideration,

$$Q_M = \frac{\mu_r''}{\mu_r'}$$

however, only when the inductance length l' is wound over the whole length of the ferrite rod. The term given in parenthesis is a coarse approximation that was found by the author experimentally.

The radiation impedance R_A of the antenna described in(3)

$$R_A = 20\,\Omega \times \left(\frac{2\pi}{\lambda_0}\right)^4 \times A^2 \times N^2$$

will become very low for large wavelengths and can usually be neglected with respect to the loss resistance R of the coil. This loss resistance results from the resistance of the wire

itself, its increase due to the skin effect, as well as the eddy current losses in the wire. The skin effect and the eddy current losses can be reduced considerably by using stranded wire, which means the DC-resistance of the wire can be considered as the loss resistance R.

The inductance L is decisive for the Q of the antenna. It is dependent, however, on the surface area A and the number of turns N, and in the case of ferrite antennas, also on the effective permeability μ_{eff}.

The inductivity L can be calculated for a round loop antenna as short, wide inductance as shown in (4):

$$L = \mu_0 \times \frac{d}{2} \times N^2 \times (\ln \frac{4d}{l} - 0.5)$$

where l is the width of the inductance and
d is the diameter of the coil.

In the case of the ferrite antenna in the form of a thin, long coil, the following is valid:

$$L = \mu_0 \frac{\pi \times d^2 \times N^2}{2\,(l+l')} \times \mu_{eff}$$

where l is the length of the rod, l' is the length of the coil, and d is the diameter of the ferrite rod. As can be seen in the diagram given in (5) shown in **Figure 3**, the effective permeability is dependent on the basic permeability μ_i of the material and the length-to-diameter l/d of the ferrite rod itself. Conventional ferrite rods for antennas having a l/d of 10 to 25, will have a μ_{eff} of between 50 and 200, which is virtually independent of the material used.

The stray capacity C_s and the input capacitance C_{in} can only be determined with the required accuracy by measuring the resonance frequency of the antenna. If the

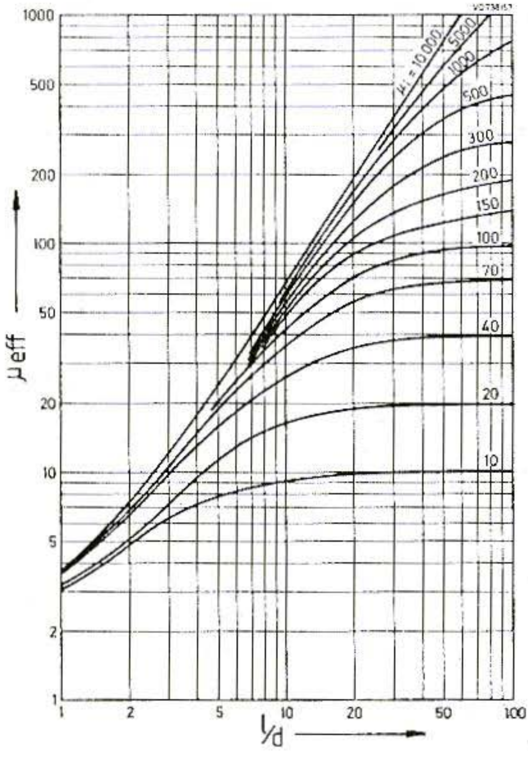

Fig. 3: Effective rod permeability μ_{eff} as a function of the l/d ratio according to (5)

antenna possesses a resonance frequency f_1 without the resonant circuit capacitor C_R, and a resonance frequency f_2 results when using a certain capacitor such as C_R = 270 pF, the following is valid:

$$C_s + C_{in} = \frac{C_R}{\left(\frac{f_1}{f_2}\right)^2 - 1}$$

This measurement allows the actual inductivity to be determined:

$$L = \frac{1}{(2\pi f_1)^2 \times (C_S + C_{in})}$$

The practical use of the equations is now to be shown with the aid of a loop and a ferrite antenna.

2.1.1. Loop Antenna

Loop diameter d = 50 cm

Coil width l = 1.3 cm

Number of turns N = 50

Stranded wire of 45 x 0.05 mm dia. silk-covered enamelled copper wire is to be used which has a resistance of 0.2 Ω/m.

Input impedance R_{in} = 220 kΩ

Permeability constant $\mu_0 = 4\pi \times 10^{-9} \frac{V\,s}{A\,cm}$

$$L = \mu_0 \times \frac{d}{2} \times N^2 \times (\ln \frac{4d}{l} - 0.5)$$

$$= 4\pi \times 10^{-9} \times \frac{50}{2} \times 50^2 \times (\ln \frac{200}{1.5} - 0.5) \frac{Vs}{A}$$

$$= 3.45 \text{ mH}$$

$$R = \pi \times d \times N \times 0.2\ \Omega/m = 15.7\ \Omega$$

If R_A is neglected, the loop antenna will have a Q of

$$Q = \frac{1}{\frac{R}{2\pi f \times L} + \frac{2\pi f L}{R_{in}}} = 58.9$$

The following was measured on the completed loop antenna:

L = 3.48 mH

$C_S + C_{in}$ = 770 pF

Q = 46.5

A capacitance value of C_R = 442 pF is required for resonating the antenna to f = 77.5 kHz. Since the stray capacitance of the winding is very dependent on the winding itself, it is necessary for C_R to be determined individually for each antenna.

The antenna voltage U_{eff} is calculated as a function of the electrical field strength E with the following measured values:

$$U_{eff} = \frac{\pi}{\sqrt{2}\ \lambda_0} \times A \times N \times Q \times \mu_{eff} \times E$$

$$= \frac{\pi}{\sqrt{2} \times 3871\text{ m}} \times \frac{\pi \times 0.5^2\text{ m}^2}{4}$$

$$\times 50 \times 46.5 \times 1 \times E$$

$$\boxed{U_{eff} = 0.26\text{ m} \times E}$$

2.1.2. The Ferrite Antenna

A ferrite rod of a length l = 200 mm and a diameter d = 10 mm is used. According to the databook, the ferrite rod has a material Q of Q_M = 180 at 77.5 kHz, and a basic permeability of μ_i = 750.

According to **Figure 3** the result is μ_{eff} = 160.

Number of turns N = 150 with a coil width l' = 30 mm, and using stranded wire of 45 x 0.05 mm dia. silk-covered enamelled wire having a resistance of R = 1.4 Ω.

If R_A is neglected, and R_{in} is assumed to be 220 kΩ, the following will result:

$$L = \mu_0 \times \frac{\pi \times d^2 \times N^2}{2(l + l')} \times \mu_{eff}$$

$$= 4\pi \times 10^{-9} \times \frac{\pi \times 1^2 \times 150^2}{2(20 + 3)} \times 160 \frac{VS}{A}$$

$$= 3.09 \text{ mH}$$

$$\text{and } Q = \frac{1}{\frac{R}{2\pi f\ L} + \frac{2\pi f\ L}{R_{in}} + \frac{3l'}{4l \times Q_M}}$$

$$= 119$$

The following was measured on the completed ferrite antenna:

$L = 2.99\ \text{mH}$

$C_S + C_{in} = 74.3\ \text{pF}$

$Q = 110$

A capacitance value of $C_R = 1.33\ \text{nF}$ is required for resonance to $f = 77.5\ \text{kHz}$.

The antenna voltage U_{eff} is calculated from the following measured values:

$$U_{eff} = \frac{\pi}{\sqrt{2} \times \lambda_0} \times A \times N \times Q \times \mu_{eff} \times E$$

$$= \frac{\pi}{\sqrt{2} \times 3871\ \text{m}} \times \frac{\pi \times (0.01)^2\ \text{m}^2}{4}$$

$$\times\ 150 \times 110 \times 160 \times E$$

$$\boxed{U_{eff} = 0.12\ \text{m} \times E}$$

It will be seen that the ferrite antenna only provides half the voltage as the loop antenna for the same field strength. This cannot be increased condiserably by increasing the number of turns N, since this will decrease the Q. When using the same ferrite rod, the following will result:

$N = 300$ turns; $Q = 32$; $U_{eff} = 0.14\ \text{m} \times E$
$N = 500$ turns; $Q = 12.5$; $U_{eff} = 0{,}145\ \text{m} \times E$

Since, on the one hand, the interference load of the input stage of the antenna amplifier increases with reducing Q, and on the other hand an interference-phase modulation occurs at higher Q, it was found that the optimum is obtained with $Q = 50$. This corresponds to a number of turns $N = 250$ in the case of a 200 mm long ferrite rod, and $N = 300$ in the case of a 140 mm ferrite rod, when using the same material and wires.

If the antenna is to be used over a wide temperature range, e.g. such as under the roof, a lower Q will be advisable due to the large temperature coefficient of the ferrite material.

3. ANTENNA PREAMPLIFIER DJ3RV 006

In order to be able to mount the antenna at a position of low interference, the same concept is used as was described in (2), where the ferrite or loop antenna was operated in conjunction with an amplifier and coupled out at low impedance using a coaxial cable.

The antenna preamplifier has thus two main tasks:
The provision of selectivity with the aid of a narrow-band filter, and the matching to the feeder cable to the receiver. The DC-voltage supply is made from the receiver via the feeder cable.

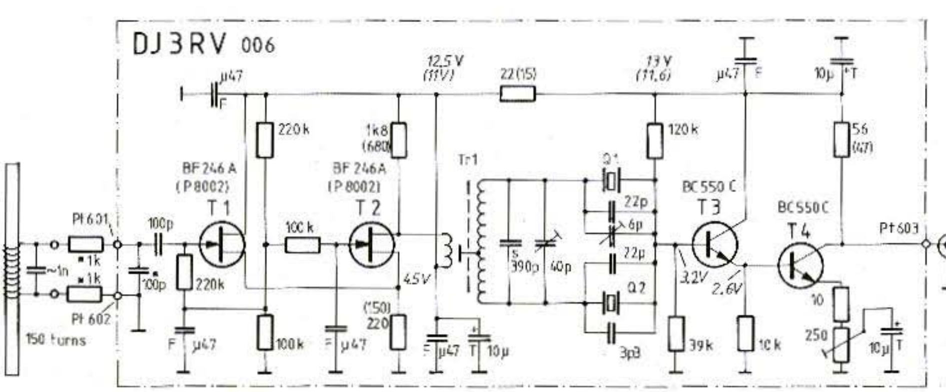

Fig. 4: Antenna preamplifier with two-pole crystal filter

3.1. Circuit Description

The circuit diagram of the module is given in **Figure 4.** In order to keep the interference load on the preamplifier as low as possible, a low and a high-pass filter are provided at the input. If the antenna is operated in the vicinity of the transmitter, it is advisable to provide the low-pass filter comprising 2 x 1 kΩ and 100 pF, , designated with * in the circuit diagram. This 100 pF capacitor is part of the resonant circuit capacitance C_R and transforms the two 1 kΩ resistances as additional loss resistances in the antenna circuit. The Q of the antenna will be reduced to Q ≈ 50 in the case of the ferrite antenna calculated in the previous section, having N = 150 turns and a circuit capacitance of approx. 1.3 nF, and a ferrite rod length of 200 mm (or approx. 1.5 nF with a length of 140 mm).

The symetrically wound transformer Tr 1 is decisive for the operation of the crystal filter. The two secondary windings must be wound at the same time and equally onto the whole core. The inductivity should amount to approx. 10 mH. Since the permeability of the cores fluctuates considerably, it is advisable to measure the coil. The number of turns can be between 2 x 27 turns, and 2 x 32 turns. When winding on or off, attention should be paid to the symmetrical characteristics of the windings. An exact alignment of the filter is otherwise no longer possible.

The high-pass filter is formed by the 100 pF coupling capacitor together with the input impedance of the amplifier, and is used for suppressing any low-frequency interference injected into the antenna.

The amplifier in front of the crystal filter comprises T 1 as source follower in order to achieve a high input impedance, and T 2 in a common gate circuit as low-reactive amplifier for decoupling the antenna and crystal filter. The 1.8 kΩ drain resistor of T 2 represents the required source impedance of the crystal filter transformed via Tr 1. The crystal filter has a bandwidth of ±55 Hz and possesses two clearly defined attenuation poles, which allow interference carriers to be suppressed by more than 70 dB.

The terminating impedance of the filter is realized with the aid of the input impedance of the emitter follower T3, and its base voltage divider. T4 is used as amplifier and possesses a low collector impedance of 56 Ω, which is used as source impedance for the feeder cable to the receiver. By altering the emitter feedback at T4, it is possible to adjust the overall gain of the module to between one and ten times to make it suitable for the various input levels.

As previously mentioned, the DC-power supply of the circuit is provided via the feeder cable. Transistors T1, T2, and T3 are fed via the 56 Ω resistor connected to the drain of T4. The filter capacitors in conjunction with the 22 Ω resistor neutralize the amplifier so that no self-oscillation can occur.

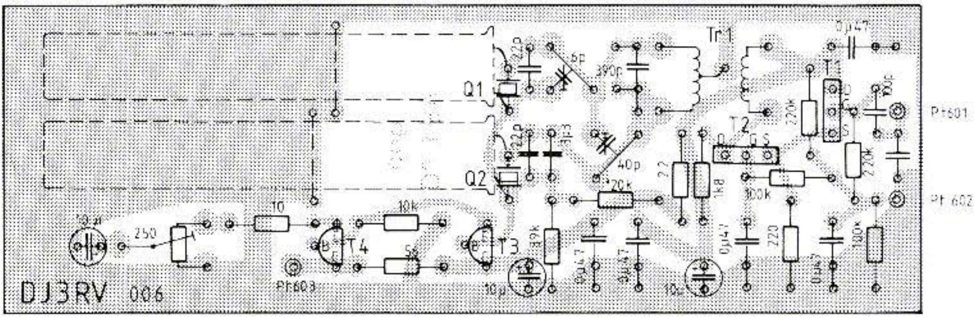

Fig. 5: Component location plan on the single-coated antenna preamplifier board DJ3RV 006

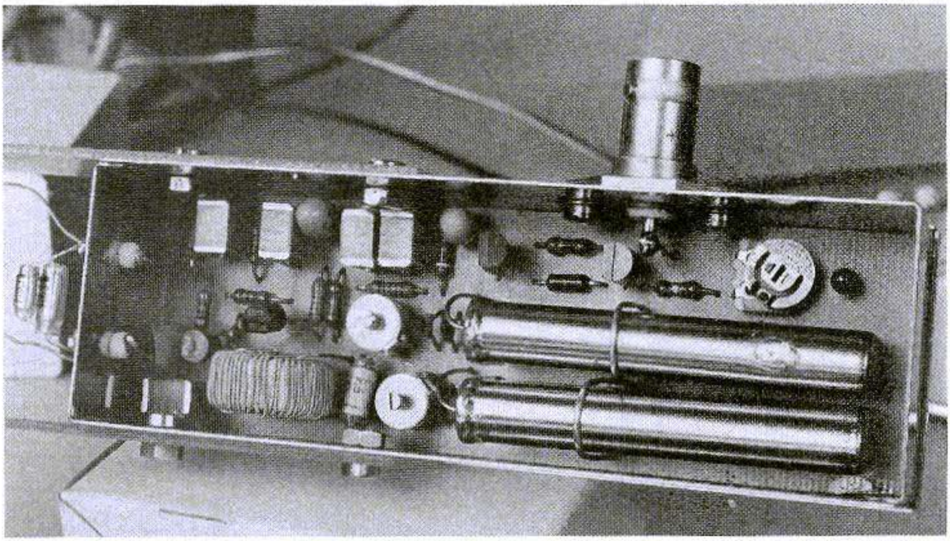

Fig. 6: Photograph of the author's prototype antenna preamplifier DJ3RV 006 complete with ferrite antenna

3.2. Construction and Alignment

The construction of the antenna preamplifier can be made without difficulties, and several DC-voltage values are given in the circuit diagram for checking the operating points. In order to ensure a balanced operating point of the input amplifier, transistors T1 and T2 should be selected to have the same U_{GS}.

The component location plan is given in **Figure 5**, and **Figure 6** shows a photograph of the author's prototype.

The following parts list contains all special components.

Parts list for DJ3RV 006

T1, T2: BF 246 A (T I)
Selected to have the same U_{GS} at I_D = 10 mA
Alternative: P 8002
Selected to have the same U_{GS} at I_D = 20 mA

T3, T4: BC 550 C (Siemens)

Tr 1: Toroid core, R 16 N 30 (Siemens)
Primary: 8 turns of stranded wire 45 x 0.05 mm dia. silk-covered enamelled copper wire
Secondary: 2 x 30 turns of stranded wire 45 x 0.05 mm dia. silk-covered enamelled copper wire

Q1: f = 77508 Hz, series resonance
L_1 = 53 H ± 20%
C_0 ≈ 10 pF

Q2: f = 77436 Hz, series resonance
L_1 = 53 H ± 20%
C_0 ≈ 10 pF

If a strong interference level is to be expected at the location of the antenna, e.g. in the vicinity of a TV-receiver or switching power supply, it is advisable to wind the antenna firstly, connect it to an oscilloscope, and align for resonance. In this case, it is possible to determine the interference level of the equipment in order to limit the number of surprises later.

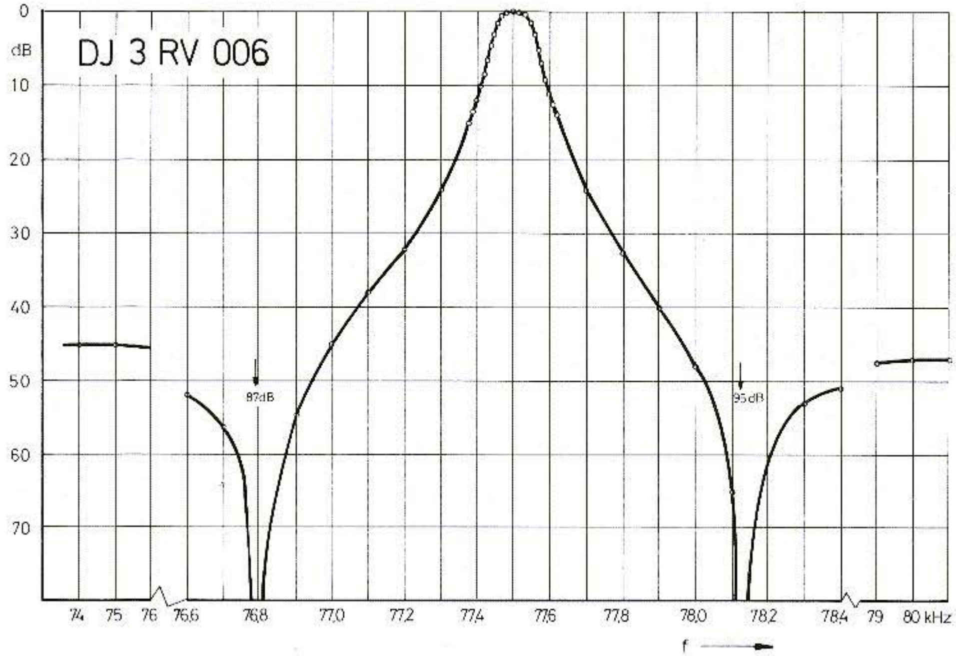

Fig. 7: The frequency response of the two-pole crystal filter used in the author's prototype

The amplifier previous to the crystal filter can be overloaded by interference pulses. At the drain of T2, limiting will take place at input voltage peaks of > 0.3 V at Pt601. If larger interference levels are to be expected, power FETs type P8002 can be used for T1 and T2. In this case, the values given in parenthesis in the circuit diagram are valid, and the primary winding of transformer Tr1 must be reduced to five turns. A limitation will then not occur until voltage peaks of > 0.8 V are exceeded.

The passband curve of the crystal filter given in **Figure 7** will result when using the crystals Q1 and Q2 listed in the parts list. The 40 pF trimmer is used to align the center frequency, and the 6 pF trimmer to adjust the position of the attenuation poles.

The alignment of the filter is made as follows:
- 6 pF trimmer to center position (approx. 4 pF)
- Align the 40 pF trimmer for maximum at 77500 Hz
- Align the 6 pF trimmer so that the attenuation pole is set to the fifth harmonic of the TV-line frequency f = 78125 Hz
- Align the 40 pF trimmer so that the attenuation at 75 kHz and 80 kHz are equal.

An alignment for the best symmetrical passband curve and ultimate selectivity can only be made with the aid of a stable generator. It is very difficult to sweep such narrow filters!

The antenna preamplifier can be used as a separate module if a coupling capacitor of at least 0.1 μF is connected to the output and the DC-voltage feed of + 12 V is fed to the collector of T 3.

4. RECEIVER AND CRYSTAL OSCILLATOR MODULE DJ3RV 007

This module is designed to obtain the time

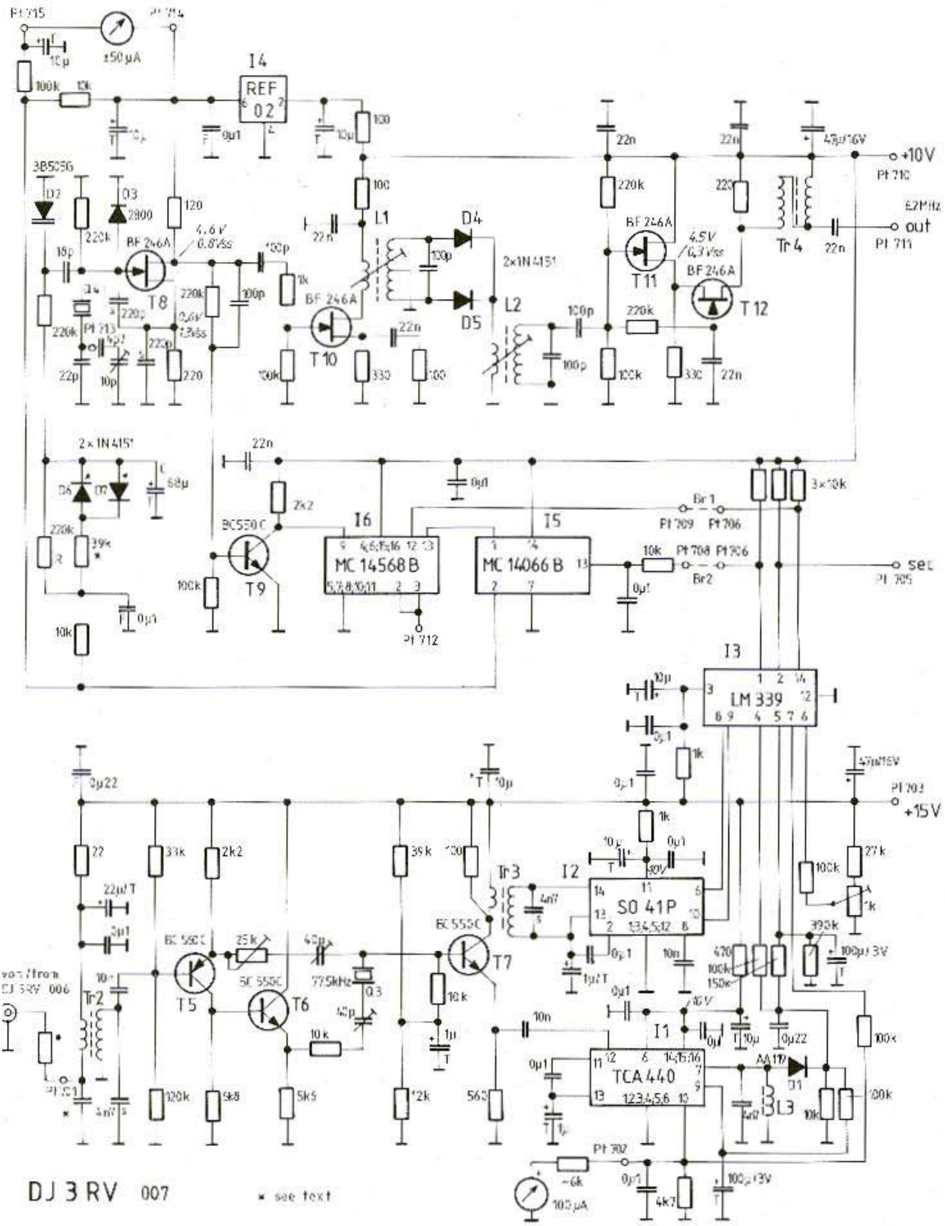

Fig. 8: Circuit diagram of module DJ3RV 007 with phaselocked crystal oscillator as frequency standard, and DCF 77 receiver with crystal filter, limiter for the reference frequency, and demodulator for the second pulses.

information and standard frequency from the DCF-77 signal provided by the antenna preamplifier. As can be seen in the block diagram given in **Figure 1**, the receiver and crystal oscillator are combined in a single module, which provides the second pulses and a 6.2 MHz signal with a level of − 6 dBm into 50 Ω.

The time information is decoded from the second pulses in a further module, which is to be described in a later edition of VHF COMMUNICATIONS. The 6.2 MHz signal is used as reference frequency for the generation of the auxiliary frequencies required for the RTTY transceiver circuits, which are also to be described later.

4.1. Circuit Description

The circuit diagram of the module is shown in **Figure 8.** The DCF 77 signal fed in from the antenna preamplifier via the coaxial cable, is fed at low impedance to Pt 701 of the toroid transformer Tr 2. At the same time, the DC-voltage supply of the antenna preamplifier is made via the primary winding. In order to improve the ultimate selectivity, the secondary winding is tuned to resonance with the aid of the 4.7 nF styroflex capacitor. This resonant circuit is very wideband due to the losses in the core material, and due to the loading of the input impedance of transistor T 5, which means that no alignment is required.

Transistors T 5 and T 6 generate the anti-phase signals for the crystal filter comprising Q 3. This part of the circuit is based on a publication of Hetzel and Rohbeck (2); the operating points and impedances were matched to the author's requirements.

Transistor T 7 isolates the filter from the subsequent stages. The limiting amplifier equipped with an SO 41 P is driven from the collector via Tr 3. The secondary winding of Tr 3 is also brought to resonance at 77.5 kHz with the aid of a 4.7 nF styroflex capacitor. The limited 77.5 kHz signal is fed via a comparator in I 3 (LM 339), which is used as level converter to CMOS-level at Pt 704. It is then passed via a bridge to Pt 709 where it is used as reference signal for phase comparison.

The DCF 77 signal is fed from the emitter of T 7 to the controlled IF-amplifier of the TCA 440. The other parts of the TCA 440 are not used, and are grounded, or connected to the operating voltage.

The amplitude-modulated second pulses are demodulated with the aid of diode D 1 (AA 119) from the amplified signal at the resonant circuit, comprising L 3 and the 4.7 nF styroflex capacitor. They are then passed via a second comparator in I 3 and converted to CMOS-level at Pt 705.

The DC-voltage from the diode is fed via a filter link comprising 100 kΩ/100 μF and used as control voltage at pin 9 of the TCA 440. Via an internal emitter follower, the control voltage is connected at low impedance to pin 10 (TCA 440), and to Pt 702. This signal is used for indicating when no transmission is being received, and is converted to CMOS-level with the aid of a third comparator in I 3 and fed to Pt 706.

A 100 μA-meter with an impedance of approx. 6 kΩ can be connected to Pt 702 for indicating the field strength.

The heart of the standard frequency generator is the low-noise crystal oscillator using a Colpitt circuit in conjunction with the junction FET T 8. The fundamental crystal Q 4 oscillates in parallel resonance and is coupled to the FET with the aid of the two 220 pF styroflex capacitors. The two capacitors should have approximately ten times the value of the alignment capacitance of the crystal. The author's prototype uses a crystal of 3.1 MHz with a parallel load of 20 pF. The frequency is aligned with the aid of the series capacitance of the crystal comprising 22 pF and 4.7 pF together with 10 pF trimmer at Pt 713. The frequency control via the PLL is made with the aid of varactor diode D 2 (BB 505 G) in series with the 18 pF capacitor at the low-impedance side of the crystal. This has several advantages. The circuit exhibits less noise, the AC-voltage amplitude at D 2 is always less than the DC-voltage amplitude, and the control slope only amounts to approx. 1 Hz/V.

If the receiver is to be often switched on and off, this control slope will be too low, and the transition time will be too long. The PC-board layout therefore provides the possibility of connecting the frequency control to Pt 713. It is necessary for the capacitance values to then be changed correspondingly.

The Schottky diode D 3 is used for amplitude limiting.

The oscillator signal is coupled out at low reaction from the drain of T 8, after which T 9 amplifies it to CMOS-level, and drives the frequency divider in I 6 (MC 14 568 B). I 6 possesses a phase comparator and two programma-

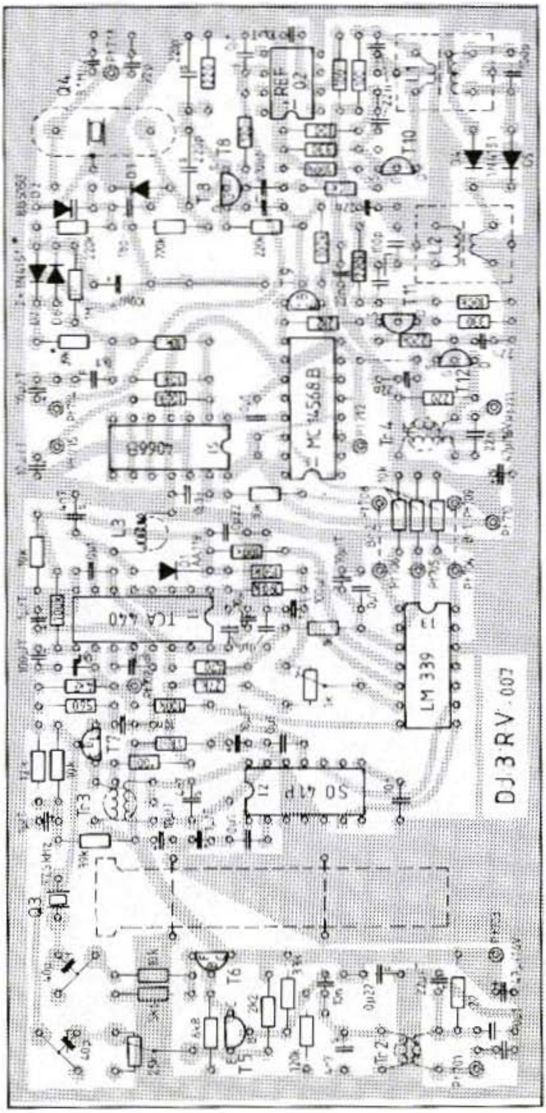

Fig. 9:
Component locations on the single-coated PC-board DJ3RV 007

ble dividers. In order to obtain a frequency division factor of 40, it is necessary for both dividers to be connected in series: The first is programmed for a frequency division factor of 1:4, and the second for 1:10.

The output of the frequency divider chain is connected to Pt 712 for alignment, and internally to the phase comparator. The reference signal from the transmitter is fed via Pt 709 and pin 12 to the phase comparator. The phase differential signal is taken from the tri-state output (pin 13) and controls the phase of the oscillator with the aid of varactor diode D 2 via CMOS switch I 5 (4066) and the low-pass filter comprising $R = 220\ k\Omega$ and $C = 68\ \mu F$.

If no signal is received from DCF 77, the control voltage will disappear at I 1, and switch I 5 will open the control circuit. Diode D 2 is then connected to the reference voltage of +5 V generated by I 4 (REF 02). The oscillator should therefore be aligned so that it has its nominal frequency with +5 V at D 2. A deviation from this value will be indicated by the meter connected between Pt 714 and Pt 715 when the control loop is closed.

The value of the low-pass filter time constant is very important. It should be as large as possible in order to reduce the effects of interference, however, be less than the instability caused by the components. This means that a large time constant requires a high-quality (preferably aged) crystal and the operation of the circuit at a virtually constant ambient temperature.

Too large a time constant will lead to an overshoot of the control circuit on switching on, or after a long period when the transmitter is not received. In the case of the described circuit, a time constant of $\tau = R \times C = 100$ s represents the upper limit. The author selected a value of $\tau = R \times C = 220\ k\Omega \times 68\ \mu F = 15$ s, which represents a useful compromise. Inspite of this, it will take approximately 3 minutes until the frequency is stable to ± 1 Hz after carrying out a 180° phase jump of the reference signal, which can be simulated by

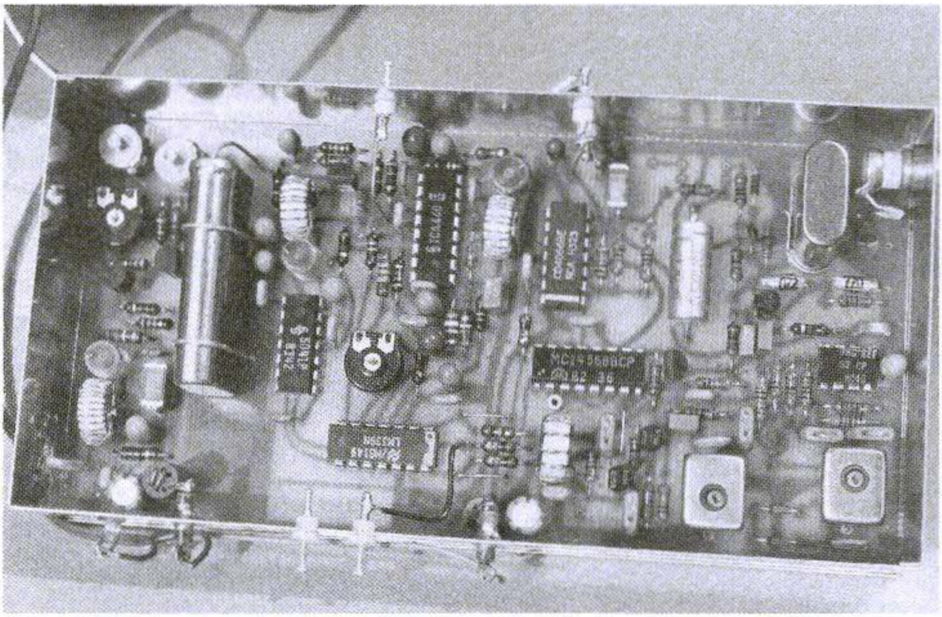

Fig. 10: Photograph of the author's prototype receiver module DJ 3 RV 007

rotating the antenna by 180°.

This time period can be reduced by connecting the series circuit of the anti-phase diodes D 6 and D 7 and the 39 kΩ resistor, designated with ✻ in the circuit diagram. These will limit large control deviations, and reduce the time constant. Since, however, any short-term phase deviations due to interference will have a large effect on the frequency stability, this possibility was not used in the author's prototype, although provided in the layout.

In order to couple out the oscillator signal with the lowest reaction possible, it is amplified with the aid of T 10 and doubled in frequency using diodes D 4 and D 5. L 1 is tuned to the crystal frequency of 3.1 MHz and L 2 to twice that frequency (6.2 MHz). The signal is amplified with the aid of T 11 and T 12 and transformed to a source impedance of 50 Ω with the aid of Tr 4.

4.2. Construction and Alignment

As can be seen in the component location plan given in **Figure 9** and the photograph of the author's prototype given in **Figure 10**, the receiver and oscillator circuit are accommodated on a common PC-board, which can be incorporated in a metal box in order to ensure that there is no danger of self-oscillation due to feedback to the antenna.

The PC-board can be completely equipped with the exception of the bridges Br 1 and Br 2. The MOS-components I 5 and I 6 should be inserted last; they can be inserted into sockets. I 1 and I 2 must be directly soldered to the board, since a tendency to self-oscillation can occur when using sockets. In order to simplify alignment of the circuit, it is advisable to measure the inductivity of the coils and the transformers before installation. The values can be taken from the following parts list:

Parts list for DJ3RV 007

T 6, T 7, T 9:	BC 550 C (Siemens, etc.)
T 5:	BC 560 C (Siemens, etc.)
T 8, T 10, T 11, T 12:	BF 246 A (TI, etc.) T 11 and T 12 selected to have the same U_{GS} at I_D = 10 mA
I 1:	TCA 440 (Siemens, etc.)
I 2:	SO 41 P (Siemens)
I 3:	LM 339 (National, etc.)
I 4:	REF 02 (PMI, Bourns)
I 5:	4066 B (RCA, etc.)
I 6:	MC 14568 B (Motorola)
D 1:	AA 119 (Siemens, etc.)
D 2:	BB 505 G (Siemens, etc.)
D 3:	2800 (HP, etc.)
D 4, D 5, D 6, D 7:	1 N 4151
Tr 2, Tr 3:	Toroid core R10 N30 (Siemens) prim.: 3.5 turns approx. 0.35 mm dia. enamelled copper wire sec.: 23.5 turns approx. 0.35 mm dia. enamelled copper wire, 0.9 mH
Tr 4:	Toroid core R10 N30 (Siemens) 2 x 12 turns, approx. 0.35 mm dia. enamelled copper wire
L 1:	Special coil kit D 41-2165, colour/orange 2 x 25 turns approx. 0.15 mm dia. enamelled copper wire, 26 μH 8 turns approx. 0.15 mm dia. enamelled copper wire
L 2:	Special coil kit D 41-2165, colour/orange 25 turns, approx. 0.15 mm dia. enamelled copper wire 7 μH 6 turns, approx. 0.15 mm dia. enamelled copper wire

L 3:	Toroid core R10 N30 (Siemens) 23.5 turns approx. 0.25 mm dia. enamelled copper wire, 0,9 mH
Q 3:	77.5 kHz, parallel 40 pF
Q 4:	3.1 MHz, parallel 20 pF
Trimmer:	40 pF plastic foil trimmer 7.5 mm dia. (Philips: violet) 10 pF air-spaced trimmer, Johanson type 5200

All other components not specifically indicated in the circuit diagram, are standard components of the given values.

Resistors: Composite carbon, spacing 10 mm, possibly metal-layer resistors in oscillator and phase-locked loop

Capacitors: Ceramic capacitors with spacings of 5 mm, some for 2.5 mm; in the oscillator circuit, use capacitors with a TC = NPO

The designations of the other capacitors are as follows:

S = Styroflex (polysterene film dielectric)
F = Plastic foil
T = Tantalum

A hermetically sealed type should be used for the low-pass filter capacitor C = 68 μF.

A signal generator, frequency counter, oscilloscope, and a multimeter are very useful for setting up and aligning the module. The following procedure is recommended by the author:

- Connect the operating voltage of + 15 V to Pt 703. The current drain should amount to approx. 25 mA without antenna preamplifier. Check the operating voltage at I 1 and I 2.
- Alignment of the crystal filter.
  Connect the signal generator via a **coupling capcitor** to Pt 701 (Attention: DC-voltage!), and connect the oscilloscope to the emitter of T 7. Align the signal generator to 77.5 kHz and select an input level of approx. 3 mV.

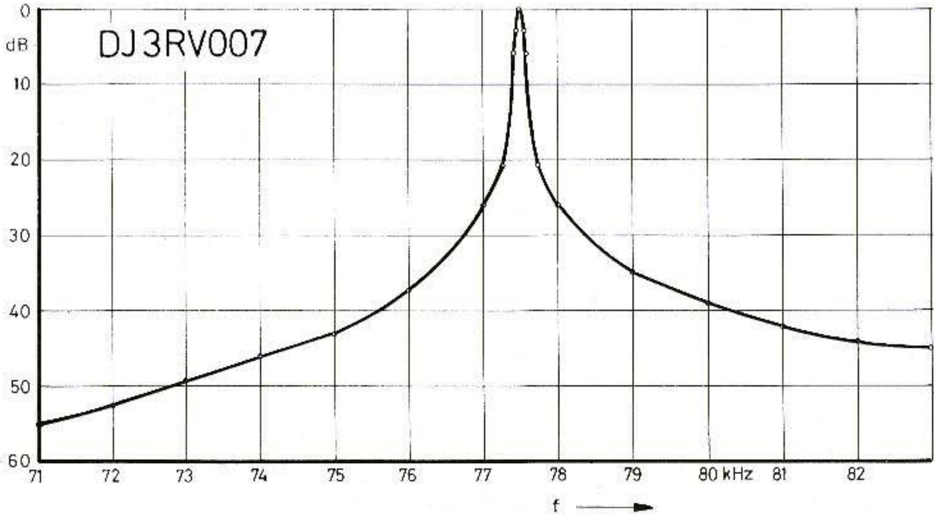

Fig. 11: Frequency response of the crystal filter in the receiver module

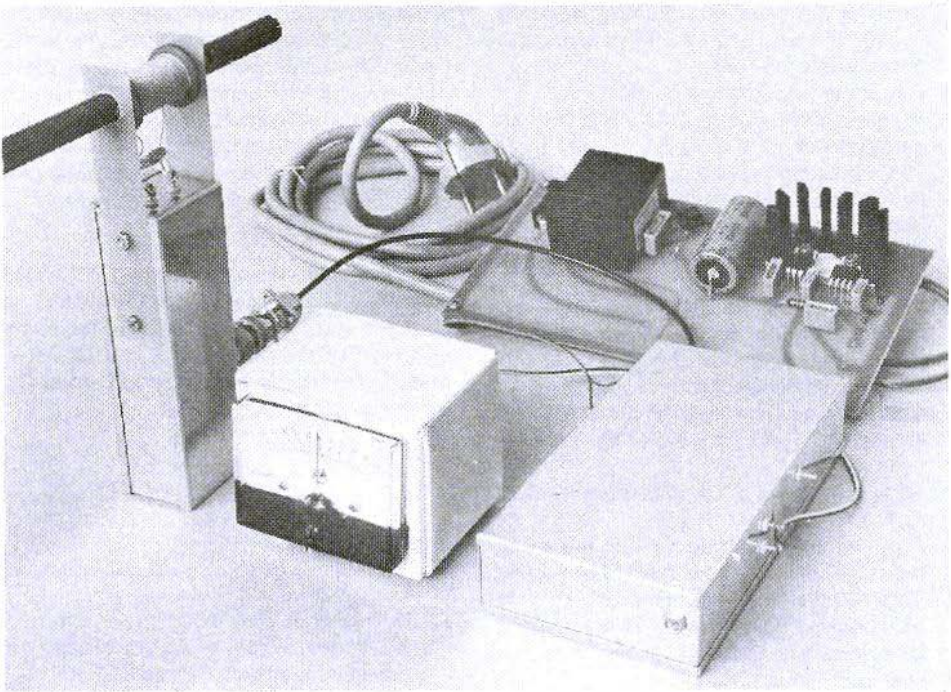

Fig. 12: The complete DCF 77 receiver ready for experimental operation

Align the trimmer in series with Q 3 for maximum level on the oscilloscope. Align the signal generator to 75 kHz, or 80 kHz with an input level of approximately 100 mV, and align the crystal filter with the bridge elements of 25 kΩ and 40 pF for best symmetrical characteristics of the ultimate selectivity. The measured frequency response of the crystal filter including the selectivity of Tr 2 is shown in **Figure 11** for the author's prototype.

- Check the receive level.

  Connect the antenna preamplifier module DJ3RV 006 to Pt 701, and point the antenna to the DCF 77 transmitter. Pay attention to any interference from the oscilloscope!

  The current drain at Pt 703 should be between 50 mA and 75 mA.

A DCF 77 signal of at least 1 mV and maximum 200 mV should be present at the emitter of T 7. A value within this range must be adjustable with the aid of the level control in the antenna amplifier. Too high a value will overdrive I 1 and I 2 and cause phase modulation of the second pulses. Too low a value will reduce the interference suppression. If the receive location is too near to the transmitter, and cannot be reduced sufficiently, it is possible for a resistor to be connected in parallel with the antenna for attenuation. It is also possible to reduce the value of the drain resistor of T 2 and simultaneously to reduce the number of turns of the primary winding of Tr 1, which will ensure a level reduction and increased interference suppression.

- Connect the operating voltage of + 10 V to Pt 710. The current drain should amount to approximately 30 mA.
  Check the following test points:
  A squarewave signal of 77.5 kHz and an amplitude of 10 V must be present at Pt 704, the second pulses at Pt 705, + 10 V at Pt 706, and + 5 V at Pt 714. The signal at Pt 706 indicates that no transmission is being received. By rotating the antenna by 90° (minimum signal), it is possible to simulate no-signal conditions. The 1 kΩ potentiometer should be adjusted so that the voltage at Pt 706 drops to approx. 0 V.
- Alignment of the oscillator circuit
  Check the voltage level at the drain and source of T 8 with the aid of the oscilloscope.
  Align L 1 and L 2 for maximum level at Pt 711.
  Align the crystal frequency with the aid of the 10 pF trimmer capacitor so that exactly 6200000 Hz can be measured at Pt 711. For checking the frequency divider in I 6, it is necessary for 77500 Hz pulses to be present at Pt 712 at CMOS-level.

  If the crystal cannot be pulled to the nominal frequency, it may be necessary for the 22 pF capacitor in series with the crystal to be increased or reduced.

- Finally, insert bridges Br 1 between Pt 704 and Pt 709, and Br 2 between Pt 706 and Pt 708. It is, of course, necessary to switch off the operating voltage and to ground the circuit and soldering iron during this process.

  After switching on the module, the PLL should synchronize the crystal oscillator to the DCF 77 signal within several minutes. This can be seen on the meter connected between Pt 714 and Pt 715 when the needle no longer jumps, but indicates a steady value.

  It may be necessary to align the frequency now and again, until the preliminary aging of the crystal and the circuit have been achieved. Any deviation from the mean value will be seen on the meter.

When using a long coaxial cable between the antenna preamplifier and receiver, it is possible for RF-signals to be injected into the cable from HF and VHF transmitters. This can be suppressed using an RC-link comprising 47 Ω and 2.2 nF connected to Pt 701. This is especially necessary when using an antenna preamplifier as described by (2).

If FETs P 8002 are used in the DJ3RV 006 module, it is necessary for the RC-link to be changed to 22 Ω and 2.2 nF, and the impedance of Tr 2 reduced from 22 to 10 Ω. Otherwise, the voltage drop of the operating voltage will be too great.

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